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Research and Development Technical Report

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UHF FIELD-EFFECT-TRANSISTOR MIXER OF HIGH DYNAMIC RANGE

BY

DAVID M. HODSDON

DECEMBER 1969

ECOM

UNITED STATES ARMY ELECTRONICS COMMAND - FORT MONMOUTH, N.J.

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Western Division
Mountain View, California 94040

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ABSTRACT

A study is presented of theoretical and experimental performance of a 90-degree-hybrid-coupled FET mixer covering 500-1000 MHz. The theoretical work completed includes methods for calculating FET mixer gain, noise figure, third-order intermodulation, and harmonic intermodulation products (single-frequency spurious responses). A model suitable for use in mixer design is also described.

Test data from the experimental model of the mixer, over the frequency range of 500 to 1000 MHz, indicated a noise figure of 10.9 to 13.3 dB, gain of 2.8 to 5.6 dB, and a 3-dB-gain-compression level of +6 dBm at the input. The intermediate frequency was 160 MHz, with a bandwidth of 4 MHz. Good correlation between calculated and measured performance was obtained. In addition, there is described a computer program, written in FORTRAN IV for the IBM 1130 and automating the calculations required to estimate FET mixer performance, in which curvative effects up through the ninth order are included.

Some comparison is made of the test results with typical hot-carrier diode mixers which, in the 500- to 1000-MHz range, are comparable except for their conversion losses. At frequencies below about 300 MHz, where the available gain of the FET mixer becomes higher and the noise figure lower, significant advantages can result from its use in a receiver system.

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Section 1

INTRODUCTION

There is a great deal of interest in field-effect-transistor (FET) mixers because they offer the possibility of low noise figure, gain, and high dynamic range. The purpose of this study was to investigate the capability of a 90-degree-hybrid-coupled FET mixer covering 500 to 1000 MHz.

This report discusses the design approach used to obtain the desired frequency coverage and test data obtained from an experimental model of the mixer. The design method used was to develop a model of the FET suitable for mixers, including noise sources, and then to imbed this model in a network which would provide performance as optimal as practicable while covering the 500- to 1000-MHz frequency range.

The performance parameters evaluated on the experimental model included gain, noise figure, input VSWR, local oscillator radiation, third-order intermodulation distortion, gain compression, and harmonic intermodulation responses resulting from undesired harmonic mixing of signal and local-oscillator frequencies. Comparisons with theoretical calculations and doubly balanced diode mixer performance are made wherever appropriate.

A computer program for calculating FET mixer performance parameters is also described. It is written in FORTRAN IV for the IBM 1130 computer, and calculates mixer constants such as conversion transconductance, single-sideband noise figure, third-order intermodulation distortion, and harmonic intermodulation responses including those caused by curvature up to the ninth order.

Section 2

FIELD-EFFECT-TRANSISTOR MODEL FOR MIXERS

In order to complete a mixer design which represents reasonable trade-offs between gain, noise figure, bandwidth and dynamic range, it is necessary to have a model which will adequately represent the performance in an actual circuit. The procedure used was to find a nonlinear model suitable for large- or small-signal analysis and then to add noise sources so that noise figures can be calculated.

The basic large-signal model (Figure 1) is considered to have an ideal square-law transfer characteristic, with a resistor in series with the source lead.[1] To account for high-frequency effects, voltage-variable capacitors are added at the gate-to-source junction and the gate-to-drain junction. The two capacitors are taken to be equal in value, since nearly all FET's are symmetrical with respect to the gate lead (i.e., interchanging drain and source leads in a circuit will not materially alter its performance). There is also a resistor in series with the drain terminal which, because of the symmetry of the device, is taken equal to the source resistance. The parameters of this model are evaluated by measuring the dc drain voltage as a function of gate-to-source voltage to obtain the transfer constants, and by measuring the low-frequency junction capacitance as a function of bias to evaluate the capacitor constants. Figure 2 shows the results of these measurements made on one of the FET's used in the mixer.

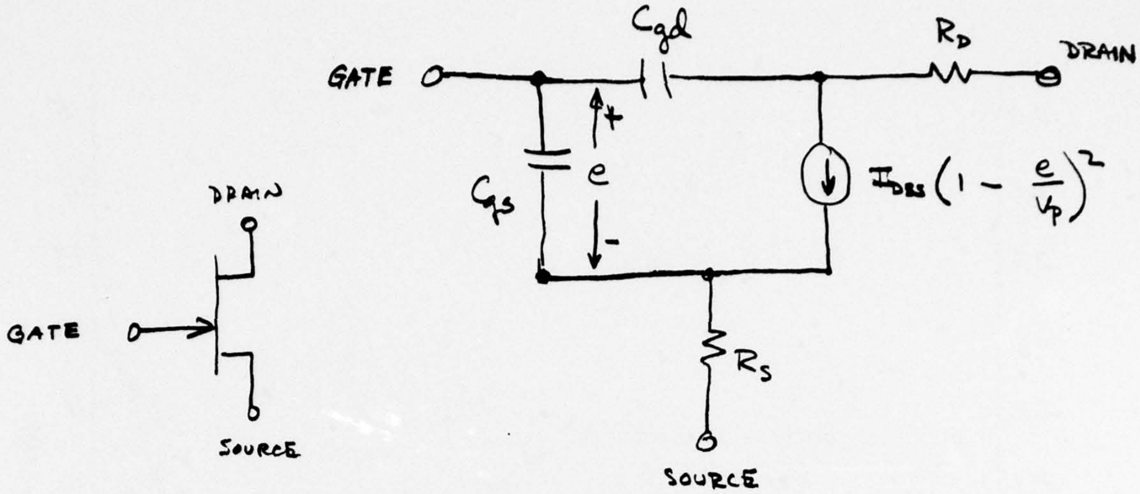
A small-signal model of the FET for a given set of bias conditions can be easily derived, using standard differentiation techniques, from the large-signal model. This has been done for one of the FET's used in the mixer (Figure 3). To check for adequacy of the model, y parameters were measured from 500 to 1000 MHz. Reasonable correlation between the computed and measured parameters was obtained (Figure 4). This indicates that a relatively simple model can be used to describe FET performance at high frequencies.

To incorporate the effect of noise into the FET model, it is necessary to identify the sources of noise. Van der Ziel[2] has shown that the primary sources of noise in an FET are bulk resistance in series with the drain and source leads, and channel-width modulation by thermal noise in the conducting channel. The former can be represented as resistors exhibiting full thermal noise, and the latter as a current generator, in parallel with the drain-to-source terminals, having the value:[2]

$$\bar{i}_n^2 = 4\,KT\,\Delta F\,\gamma\,g_{max}\ ,$$

where:
$$g_{max} = \frac{2\,I_{DSS}}{V_p}\left(1 - \frac{e}{V_p}\right) \qquad (1)$$

[1]Superscript numerals in the text denote references listed in Section 7.



$$C_{gc} = C_{gd} = \frac{C_0}{\left(1+\frac{d}{V}\right)^{\alpha}}$$

$$C_{gs} = \frac{C_0}{(d+V_{gs})^{\alpha}}$$

$$C_{gd} = \frac{C_0}{(d+V_{gd})^{\alpha}}$$

α = LAW OF CAPACITANCE VARIATION FOR JUNCTIONS

d = WORK VOLTAGE FOR JUNCTION, USUALLY 0.75 VOLTS.

C_0 = JUNCTION CAPACITANCE AT ZERO VOLTS

I_{DSS} = DRAIN CURRENT FOR $e = 0$.

V_P = GATE-TO-SOURCE VOLTAGE FOR DRAIN-CURRENT EQUAL TO ONE MICROAMP

Figure 1. FET Large-Signal Model.

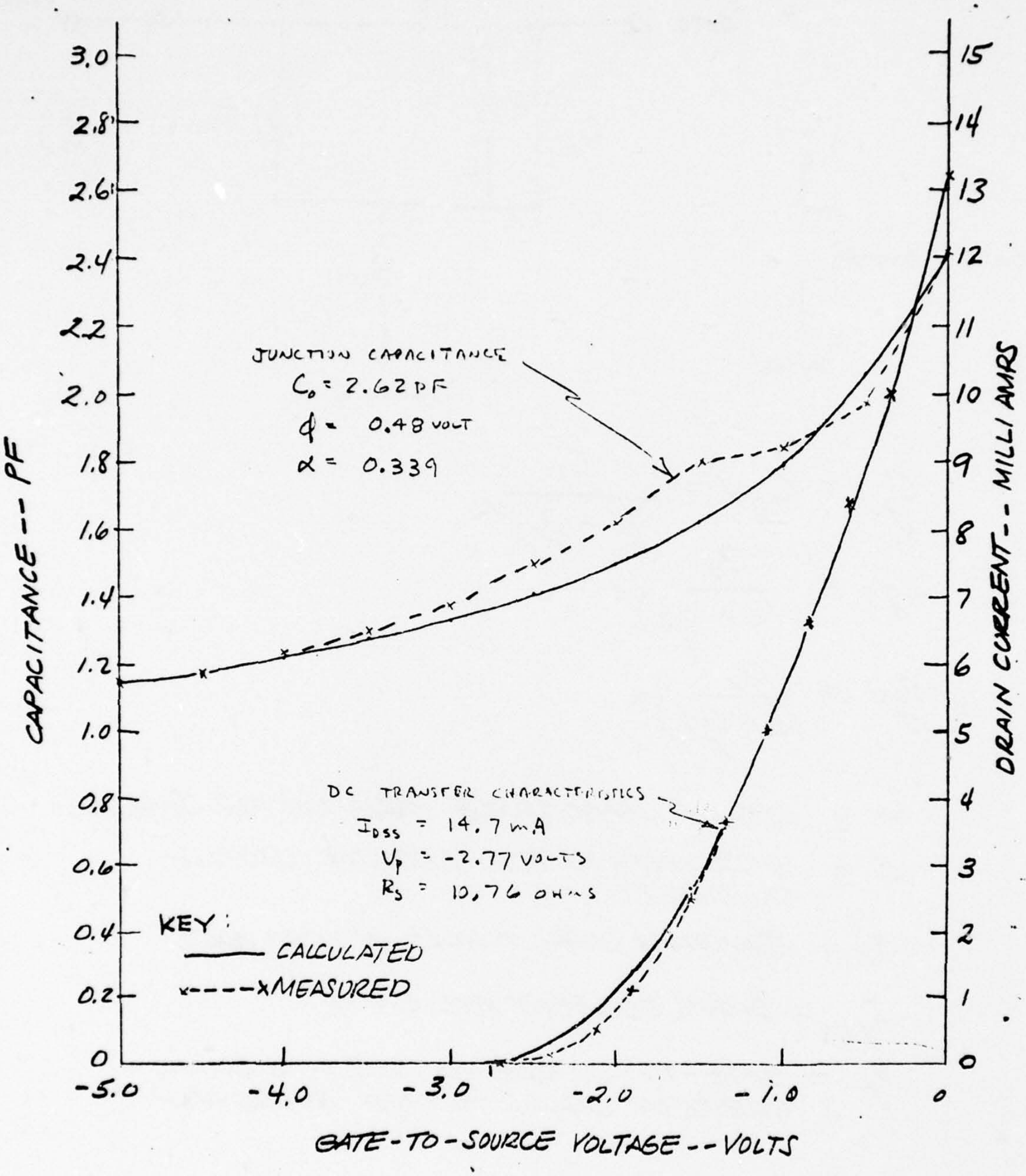


Figure 2. Comparison of Measured and Calculated Characteristics of a Field-Effect Transistor.

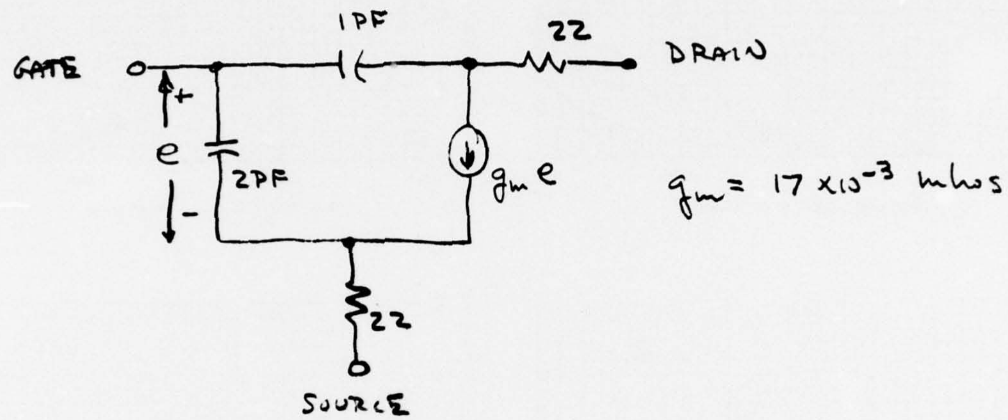


Figure 3. Small-Signal Equivalent Circuit for a High-Frequency Field-Effect Transistor.

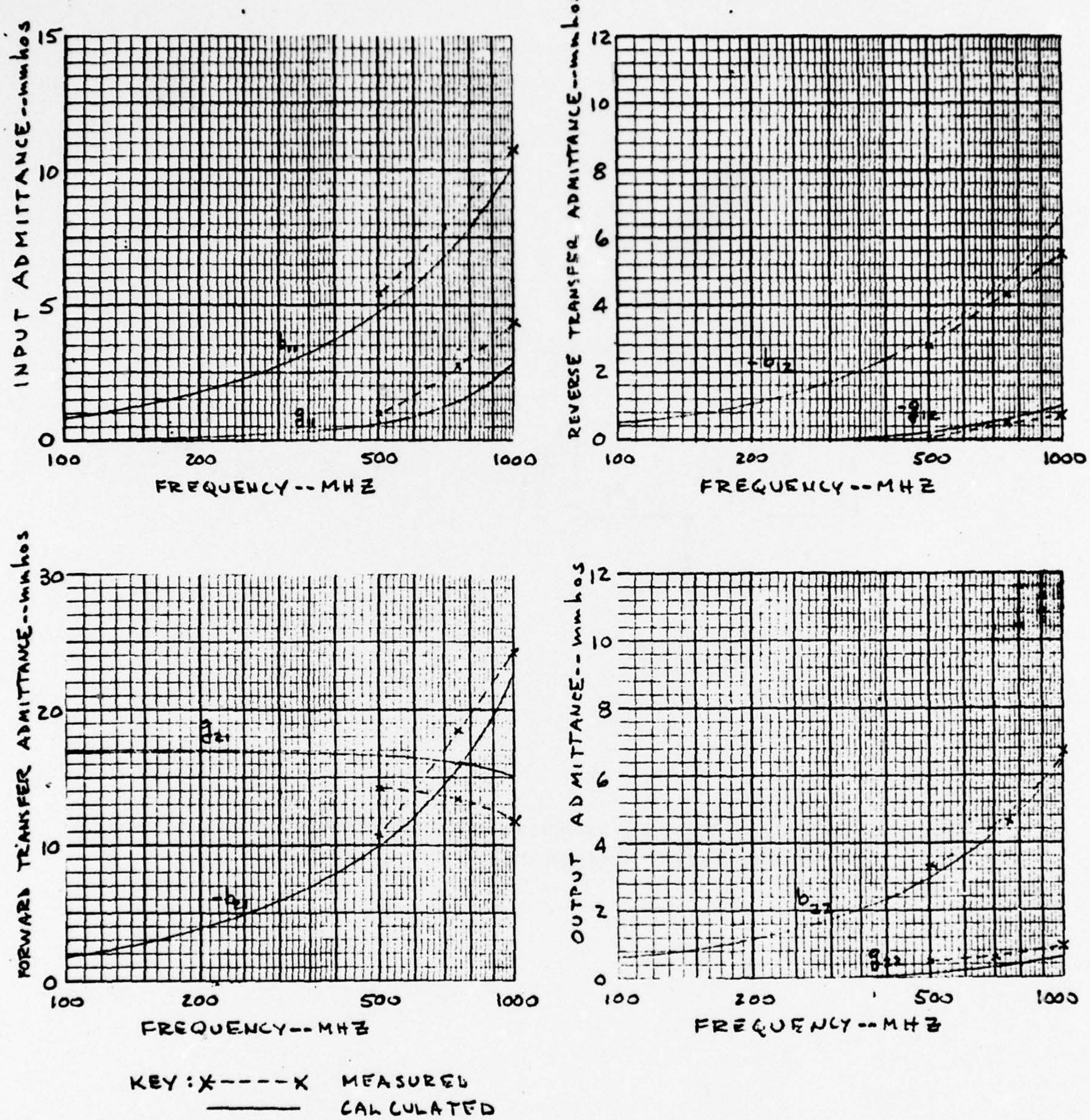


Figure 4. Comparison Between Computed and Measured 'Y' Parameters for a UHF Field-Effect Transistor -- I_D = 10 ma, V_{DS} = 10 Volts.

γ = 0.6

K = Boltzmann's constant

T = absolute temperature -- kelvin

ΔF = measurement bandwidth -- hertz

Figure 5 illustrates the complete model, including noises.

The model developed thus far includes the major effects found in FET's and can be used for ac analysis. It is not suitable for dc analysis where the input conductance at the gate terminal is important. This resistance is of the order of 10 to 20 megohms for most FET's. In addition, the FET's considered are junction types and the model does not apply if the gate-to-source junction is forward-biased by more than about 0.5 volt (in the junction FET a diode is formed from the gate-to-source and is reverse-biased for normal operation).

To develop a small-signal model suitable for use in a mixer, it is necessary to consider the effect of a time-varying local-oscillator voltage applied to the gate terminal. This voltage is taken to be sinusoidal; the circuit to be considered is shown in Figure 6. Two cases of local-oscillator drive are to be considered. The first is the square-law case wherein

$$\mathrm{VLO} + \mathrm{VGO} \leq \mathrm{VP} \text{ (square-law mixing).} \tag{2}$$

The other is the large-signal case wherein excursions into the pinch-off region are permitted,

$$\mathrm{VLO} + \mathrm{VGO} > \mathrm{VP} \text{ (large-signal mixing).} \tag{3}$$

If the conversion transconductance GC is defined as

$$\mathrm{GC} = \frac{\text{Drain Current at } \omega_{IF}}{\text{Gate-to-Source Voltage at } \omega_R} = \frac{i_D(\omega_{IF})}{e(\omega_R)} \tag{4}$$

and the average transconductance, GM is defined as

$$\mathrm{GM} = \frac{\text{Drain Current at } \omega_R}{\text{Gate-to-Source Voltage at } \omega_R} = \frac{i_D(\omega_R)}{e(\omega_R)}\ , \tag{5}$$

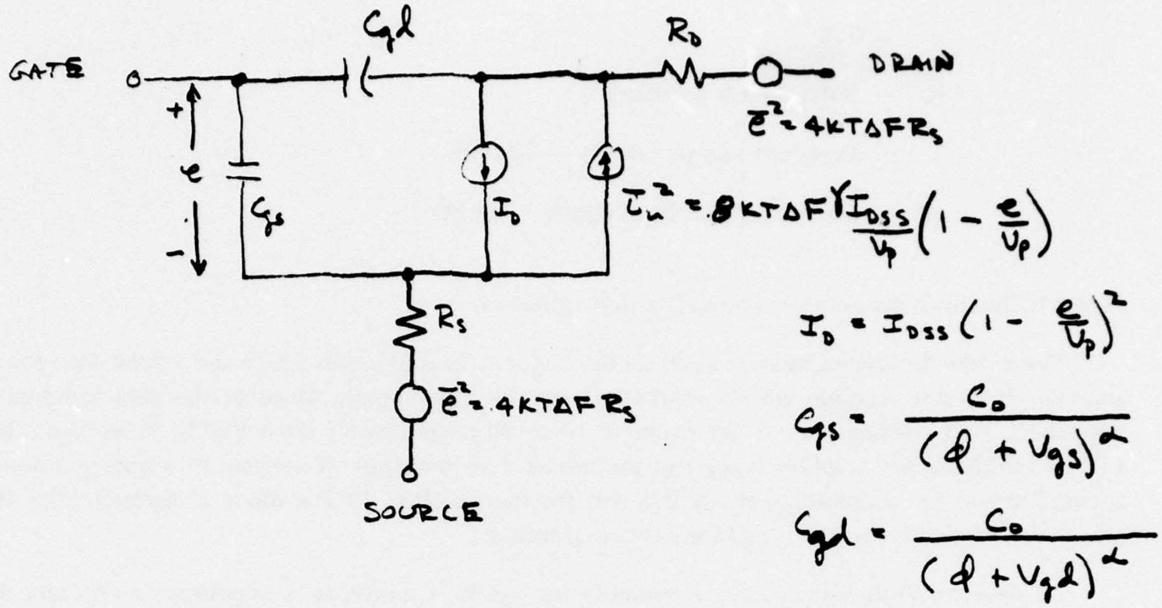


Figure 5. FET Large-Signal Model, Including Noise Sources.

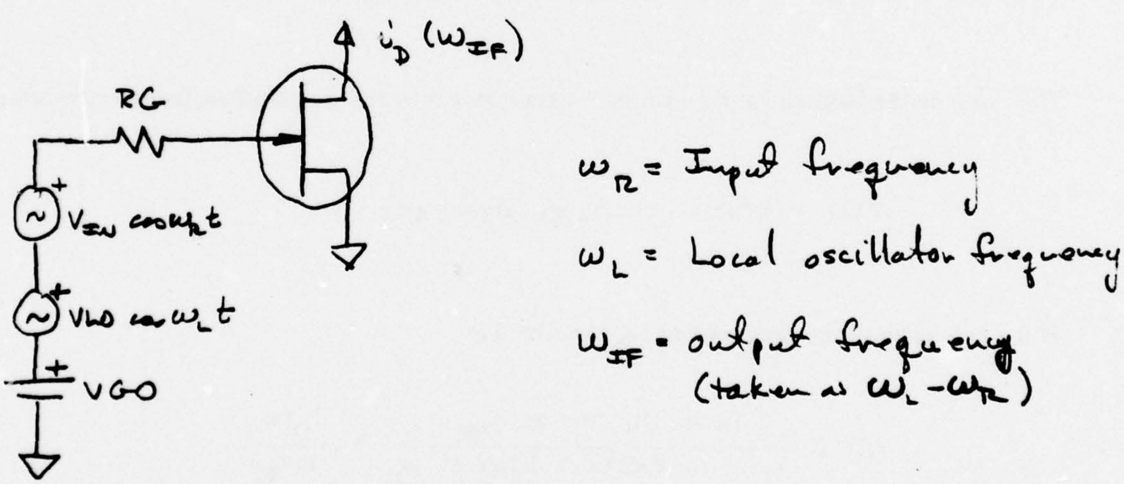


Figure 6. FET Mixer Circuit.

then it can be shown, in the square-law case, that[3]

$$GC = \frac{I_{DSS}}{VP^2} \; VLO \quad , \tag{6}$$

and that

$$GM = \frac{2\, I_{DSS}}{V_p} \left(1 - \frac{VGO}{V_p} \right). \tag{7}$$

For the large-signal case, the effect of excursions of the local oscillator into the pinch-off region must be considered. Since the square-law representation of the FET transfer characteristic is not valid beyond pinch-off, it is necessary to consider the Fourier series representation of a fractional sine wave for the local oscillator voltage, as shown in Figure 7. Comparing this with the general Fourier series representation given in Figure 8, the following equivalences are given:

$$A = VLO + (VP - VGO) \tag{8}$$

$$v\,(t_o/2) = VGO - VLO \cos\,(\pi\, t_o/T) \tag{9}$$

$$v\,(t_o/2) = VP \; . \tag{10}$$

Equating (7) and (8) and solving for $\cos\,(\pi t_o/T)$ yields

$$\cos\left(\pi \frac{t_o}{T}\right) = \frac{VGO - VP}{VLO} \quad \text{for } 0 \le \pi \frac{t_o}{T} < \pi \tag{11}$$

or

$$\frac{\pi\, t_o}{T} = \cos^{-1}\left(\frac{VGO - VP}{VLO}\right) \; .$$

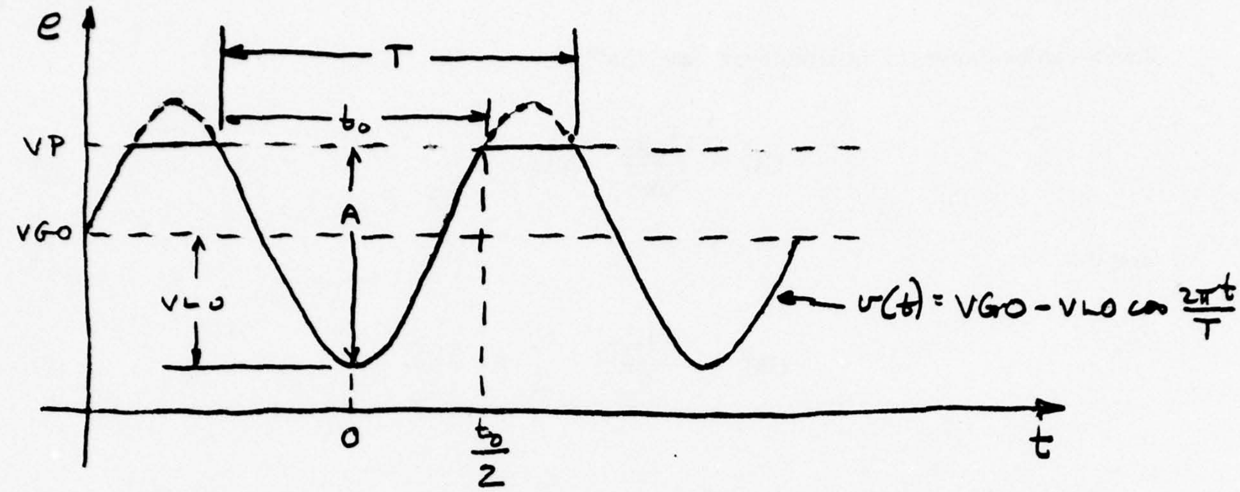
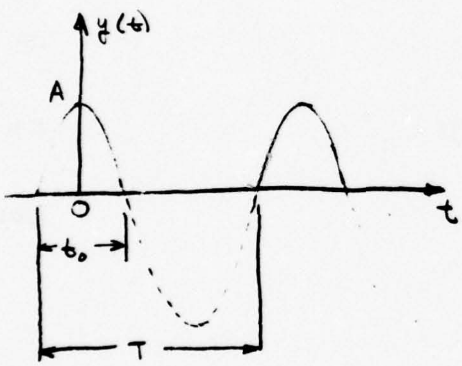


Figure 7. Representation of Gate-to-Source Voltage with a Local-Oscillator and dc Voltage Applied.



$$\alpha = \frac{\pi t_o}{T}$$

$$A_{AV} = \frac{A}{\pi} \cdot \frac{(\sin\alpha - \alpha\cos\alpha)}{1 - \cos\alpha}$$

$$C_1 = \frac{A_{AV}(\alpha - \frac{1}{2}\sin 2\alpha)}{(\sin\alpha - \alpha\cos\alpha)}$$

$$C_N = 2A_{AV}\left|\frac{\sin N\alpha\cos\alpha - N\sin\alpha\cos N\alpha}{N(N^2-1)(\sin\alpha - \alpha\cos\alpha)}\right|$$

$$y(t) = A_{AV} + C_1\cos\frac{2\pi t}{T} + \sum_{N=2}^{\infty} C_N\cos\frac{2N\pi t}{T}$$

Figure 8. Fourier Series Representation of a Fractional Sine Wave.[4]

If the substitution $a = \pi t_o/T$ is made, then the Fourier series representation of the gate-to-source voltage for the large-signal case is given by:

$$v(t) = VP - \frac{(VLO + VP - VGO)(\sin\alpha - a\cos\alpha)}{\pi(1-\cos\alpha)}$$

$$\left[1 + \frac{(\alpha - 0.5\sin 2\alpha)}{\sin\alpha - \alpha\cos\alpha}\cos\frac{2\pi}{T}t \right. \qquad (12)$$

$$\left. + \sum_{N=2}^{\infty} \frac{\sin N\alpha\cos\alpha - N\sin\alpha\cos N\alpha}{N(N^2-1)(\sin\alpha - \alpha\cos\alpha)}\cos\frac{2N\pi}{T}t_o\right]$$

where $\alpha = \cos^{-1}\left(\frac{VGO - VP}{VLO}\right); \quad 0 \leq \alpha < \pi$

$$\frac{2\pi}{T} = \omega_L .$$

With equation (12) as a representation of the local oscillator voltage, it can then be shown that:*

$$g_c = \frac{I_{DSS}}{V_p^2}\frac{(VLO + VP - VGO)(\alpha - 0.5\sin 2\alpha)}{\pi(1-\cos\alpha)} \qquad (13)$$

and that

$$g_n = \frac{2I_{DSS}}{V_p}\left(1 - \frac{VGOP}{V_p}\right)$$

*This is equivalent to the equation for g_c given in Reference 3, p 6, Equation (4).

where $$\mathrm{VGOP} = \frac{\mathrm{VP} - (\mathrm{VLO} + \mathrm{VP} - \mathrm{VGO})(\sin\alpha - \alpha\cos\alpha)}{\pi(1 - \cos\alpha)} \tag{14}$$

The FET mixer is now defined, with the exception of the change in the noise sources in the device caused by the local-oscillator voltage. It can be seen from Equation (1) that the noise in the channel is dependent upon the gate-to-source voltage. To determine the effective noise current in the channel, the large-signal case will be considered first, and then the square-law case.

The channel noise is, in fact, thermal noise; but since ΔF is usually small, it is sufficient for analysis purposes to consider it sinusoidal. Thus, let:

$$\sqrt{\overline{i_n^2}} = i_n = i_o \left(1 - \frac{e}{VP}\right)^{1/2} \cos \omega t \ , \tag{15}$$

where $$i_o = \sqrt{\frac{8 K T \Delta F \gamma I_{DSS}}{VP}}$$

ω = some frequency of interest.

The primary noise frequencies of interest are ω_{IF}, ω_R, and $\omega_L + \omega_{IF}$, the image frequency. With this in mind, the total noise current is

$$i_n = i_o \left(1 - \frac{e}{VP}\right)^{1/2} \left(\cos \omega_{IF} t + \cos \omega_R t + \cos (V_L + \omega_{IF}) t\right) . \tag{16}$$

Using the binomial expansion, we get

$$\left(1 - \frac{e}{VP}\right)^{1/2} = 1 - \frac{e}{2\,VP} - \frac{e^2}{8\,VP^2} - \frac{e^3}{16\,VP^3} - \cdots \tag{17}$$

Let

$$e = V(t) = VGOP + V1 \cos \omega_L t + V2 \cos 2\omega_L t + V3 \cos 3\omega_L t + V4 \cos 4\omega_L t + \ldots, \tag{18}$$

where V1, V2, V3, V4 are defined by Equation (12).

Now, substituting (18) into (17) and then (17) into (16), it is possible to show, after collecting like terms and using the trigonometric identity $\cos x \cos y = (1/2) \cos (x + y) + (1/2) \cos (x - y)$, that the channel noise current at the output frequency is:

$$\begin{aligned} i_n(\omega_{IF}) = \; & i_o \left[\left(1 - \frac{VGOP}{2\,VP} - \frac{VGOP^2}{8\,VP^2} - \ldots \right) \cos \omega_{IF} t \right. \\ & - \frac{V1}{4\,VP}\left(1 + \frac{VGOP}{2} + \frac{V2}{4\,VP} + \frac{V2V3}{4\,V1VP} + \frac{V3V4}{4V1VP}\right) \cos (\omega_L - \omega_R)\, t \\ & \left. - \frac{V1}{4\,VP}\left(1 + \frac{VGOP}{2} + \frac{V2}{4VP} + \frac{V2V3}{4V1VP} + \frac{V3V4}{4V1VP}\right) \cos (\omega_L + \omega_{IF} - \omega_L)\, t \right]. \end{aligned} \tag{19}$$

The first term is just the average noise current at the output frequency, since

$$\left(1 - \frac{VGOP}{VP}\right)^{1/2} = 1 - \frac{VGOP}{2VP} - \frac{VGOP^2}{8VP^2} - \ldots . \tag{20}$$

The second and third terms derive from noise current at the image and input frequencies which has been converted to the output frequency.

Since the three components of noise are uncorrelated, it is necessary to add them in an rms sense. Hence, the total output noise current squared, using Equations (19), (20) and (15), is given by

$$\overline{i_n(\omega_{IF})^2} = \frac{8\,K\,T\,\Delta F\,\gamma\, I_{DSS}}{VP} \left\{ \left(1 - \frac{VGOP}{VP}\right) + 2\left[\frac{V1}{4VP}\left(1 + \frac{VGOP}{2} + \frac{V2}{4VP} + \frac{V2V3}{4V1VP} + \frac{V3V4}{4V1VP}\right)\right]^2 \right\} \tag{21}$$

(large-signal case) .

The results for the square-law case can be obtained by letting VGOP = VGO, V1 = VLO, V2 = V3 = V4 = 0.

Thus,

$$\overline{i_{n\,\omega_{IF}}}^{\,2} = \frac{8KT\Delta F\gamma I_{DSS}}{VP}\left[\left(1 - \frac{VGO}{VP}\right) + \frac{\overline{V1}^{\,2}}{2VP^2}\right] \qquad (22)$$

(square-law case).

This completes the derivation of the FET mixer model. The final model is shown in Figure 9. This model was used to analyze the mixer design to follow, and as a basis for the computer program described in Section 5.

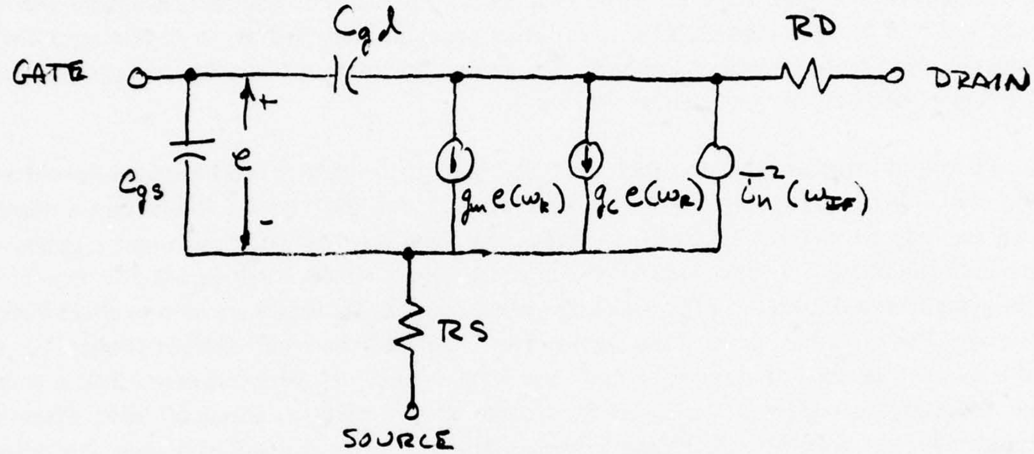


Figure 9. FET Small-Signal Mixer Model.

Section 3

MIXER DESIGN

The design approach selected was to use a 3-dB, 90-degree hybrid[5] to couple the signal and local-oscillator frequencies into two field-effect transistors preceded by a band-pass matching network. Since this results in currents at the drain of the two FET's, which are 180 degrees out of phase, a tuned transformer is employed to recombine the two outputs. A block diagram of the mixer is shown in Figure 10. The hybrid coupler is a standard commercially available unit designed to operate in a 50-ohm system. Thus, the design problem consisted of, first, finding a network to match 50 ohms to the input of the FET and, second, generating a suitable transformer design for the intermediate-frequency output of the mixer.

The input impedance of the FET from 500 MHz to 1000 MHz can be approximated as 3 pf in shunt with 300 ohms. Using the results of Plotkin and Nahi,[6] it was found that a mismatch of 3-to-1 in impedance was required to cover the 500- to 1000-MHz frequency range, using a three-pole matching network. More poles in the network would result in a better match, but the improvement was not sufficient to make it worthwhile. Calculations were then made at 750 MHz to determine the probable mixer noise figure. The results of these calculations (Figure 11) indicate that a source resistance of about 400 ohms would be optimal. In order to achieve this, it would have been necessary to reduce the frequency coverage of the mixer to about 50 MHz. Since this was substantially less than the 500-MHz coverage desired, it was elected to accept the noise figure resulting from the 100-ohm source.

Preliminary calculations indicated that 6 dB of mixer gain could be obtained with a 5000-ohm load to the intermediate frequency at the drain of the FET. Such a high resistance is required because of the low conversion transconductance (3 millimhos) and the relatively low source resistance. In addition, 5 to 6 pf of shunt capacitance is required at the drain terminal of each FET to adequately bypass the input and local-oscillator signals. With this load resistance and shunt capacitance in mind, a standard double-tuned transformer[6] was designed for the output network at the intermediate frequency of 160 MHz. The resulting bandwidth of the output circuit was about 4 MHz. This bandwidth is narrower than desired in many cases, but was necessary because of the high load-resistance and shunt-capacitance requirements.

With the circuit design complete (Figure 12) a detailed analysis of its performance was undertaken. The technique used in analyzing the FET mixer for noise figure was to accurately model the major sources of noise in the mixer and to then compute the circuit noise figure, taking into account noise contributions at the input, image and intermediate frequencies. The noise sources considered were those of the FET and losses in the input and output matching networks. It is possible to develop mathematical expressions for the noise figure, as is done for amplifiers, but because of the three frequencies involved the formulas become completely unwieldy, except for the

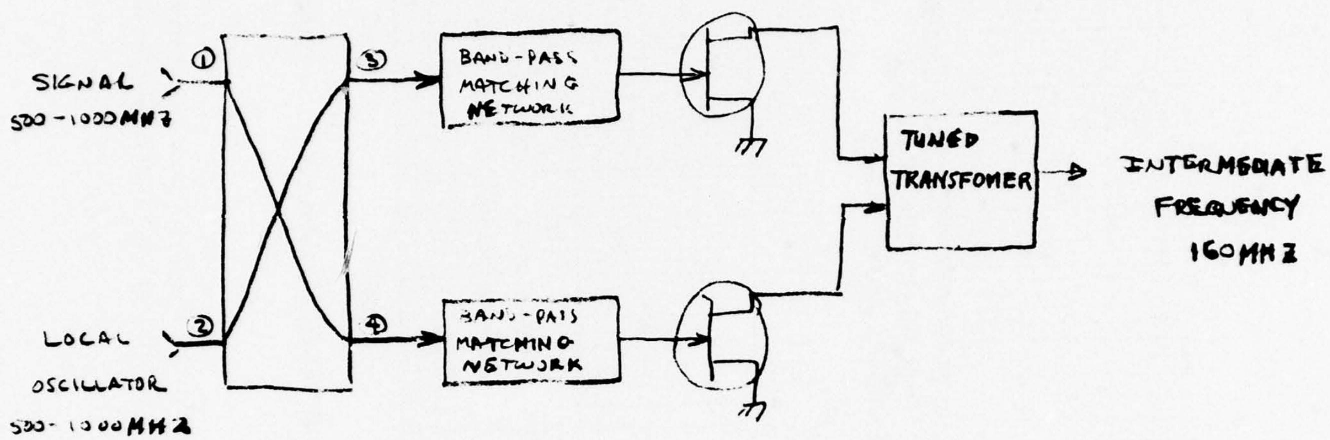


Figure 10. UHF Field-Effect Transistor Mixer -- Block Diagram.

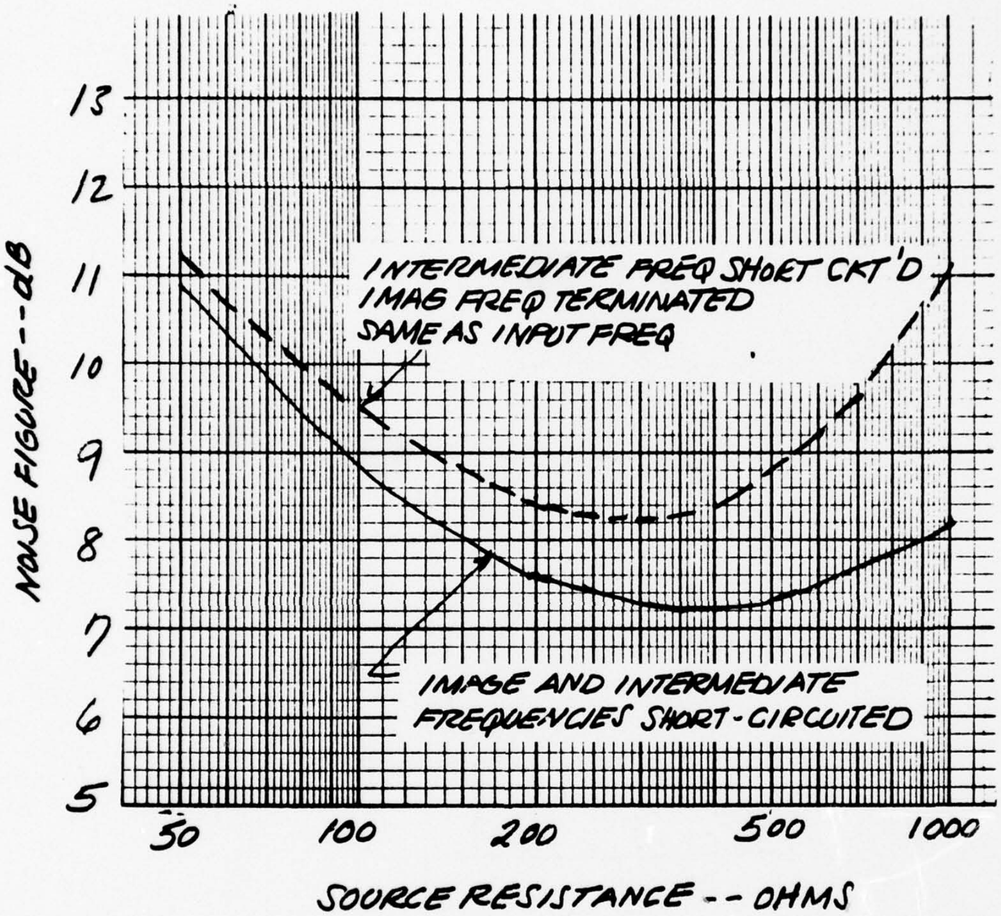


Figure 11. Calculated Noise Figure for a 750-MHz Field-Effect Mixer as a Function of Source Resistance -- IF = 160 MHz.

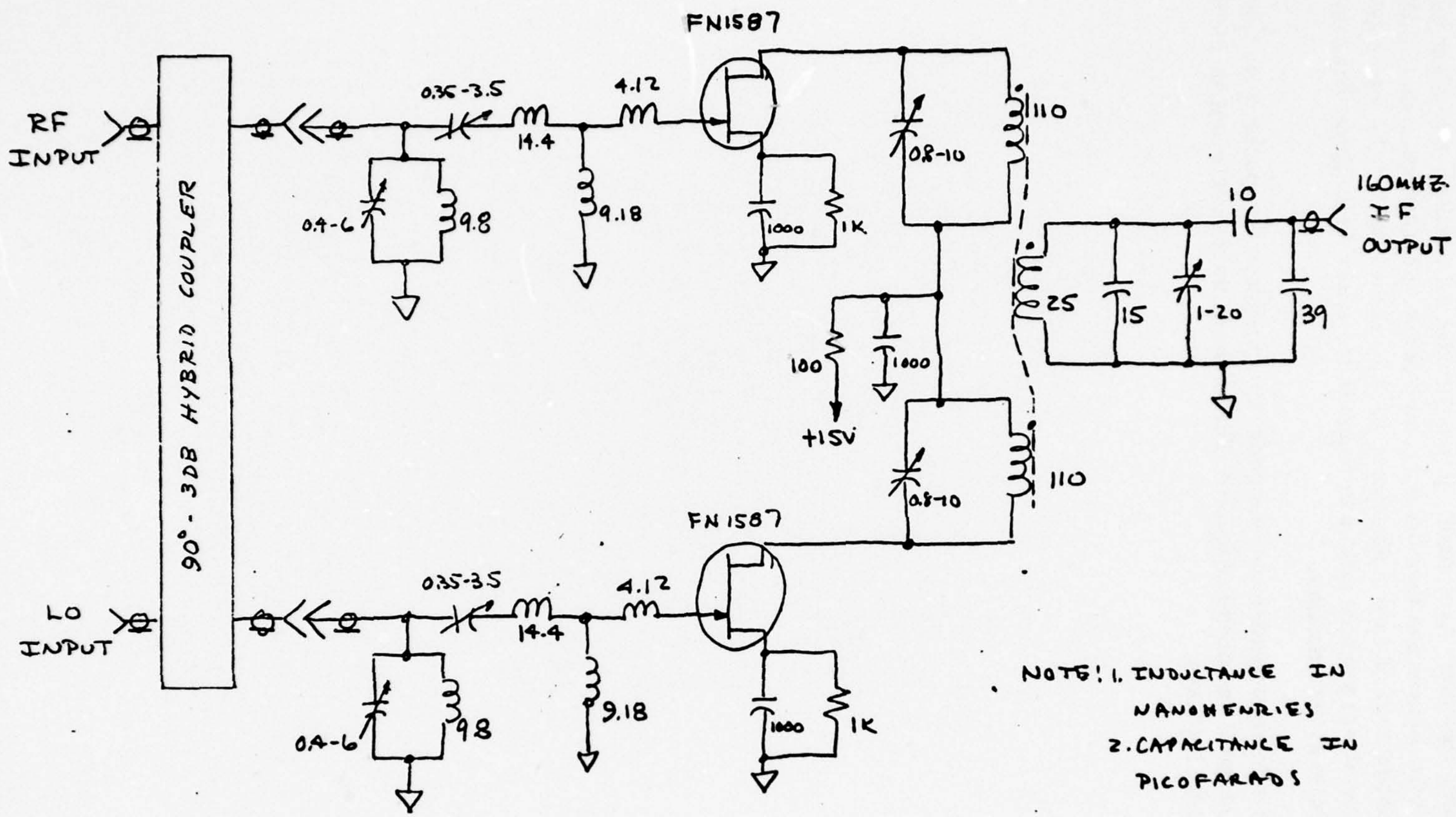


Figure 12. Circuit Diagram for a UHF Field-Effect Transistor Mixer Covering 500 to 1000 MHz.

most elementary structure. As a result, it was decided to calculate the mixer noise figure by modeling the noise sources and then using a computer circuit-analysis program (ECAP), thus determining the contribution of each noise source to the output noise of the mixer. With this information available, it was a simple matter to compute the mixer noise figure. The gain of the mixer was also calculated at the same time.

Third-order intermodulation and harmonic intermodulation responses were calculated, using the computer program described in Section 5. The results of these calculations are included with the performance data, Section 4.

Section 4

PERFORMANCE DATA FROM EXPERIMENTAL MODELS

Two mixers were constructed as a part of this program. One was a 90-degree-hybrid-coupled experimental model of the design discussed in Section 3. The other was built at 10 MHz to check the soundness of the FET model discussed in Section 2.

The experimental model of the 90-degree-hybrid-coupled mixer (Figure 13) was built by using microstrip techniques. The test data indicate a frequency coverage of 500 MHz to 1000 MHz, with a noise figure of from 10.9 to 13.3 dB, gain of 2.8 to 5.6 dB, and a 3-dB-gain-compression level of +6 dBm input. The overall performance compares with calculated values to within 2 or 3 dB, with the exception of intermodulation responses. Noise figure, gain, saturation, input VSWR, local oscillator radiation, and third-order and harmonic intermodulation performance were measured, with results indicated in Figures 14, 15 and 16 and Tables 1 and 2.

Comparisons between a good commercially available, double balanced, hot-carrier diode mixer and the FET mixer are made wherever appropriate. It is estimated from the comparison that presently available diode mixers operating in the 500- to 1000-MHz range will provide performance comparable to that of the FET mixer developed during this program. The data presented in the test results for doubly balanced diode mixers were taken from typical data given for the Relcom M1 mixer. These data were taken at 50 MHz. It is assumed that a properly optimized diode mixer operating in the 50- to 1000-MHz range would give similar performance. Discussions with Relcom in Mountain View, California, indicate that their published data on the single-frequency spurious output of the M1 mixer are applicable to higher-frequency units such as the M1A (500 to 1000 MHz) provided the RF and LO input signal frequencies are in the vicinity of 100 MHz. Above 100 MHz, some degradation can be expected because of poorer balance and poorer isolation from high-order harmonics having frequencies above the upper limit of the operating range (above 1000 MHz).

One of the aims in constructing the mixer was to compare calculated and measured performance so as to obtain a measure of the model. The test results showed that the gain and noise-figure calculations were reasonable, but additional work on the model for intermodulation performance is required.

It is felt that noise-figure and gain discrepancies between calculated and actual performance result from realization problems. For example, during alignment of the mixer, it became apparent that phase-matching between halves of the mixer affected noise figures significantly. In fact, the peaks at 600 MHz and 750 MHz can be reduced from about 13 dB to 11 dB with very small adjustments of phase balance between the IF transformers. Differences between measured and calculated values of third-order and harmonic intermodulation responses can be attributed to inadequacies in the model of the field-effect transistor used. More work is required on the model to improve correlation between predicted and actual performance.

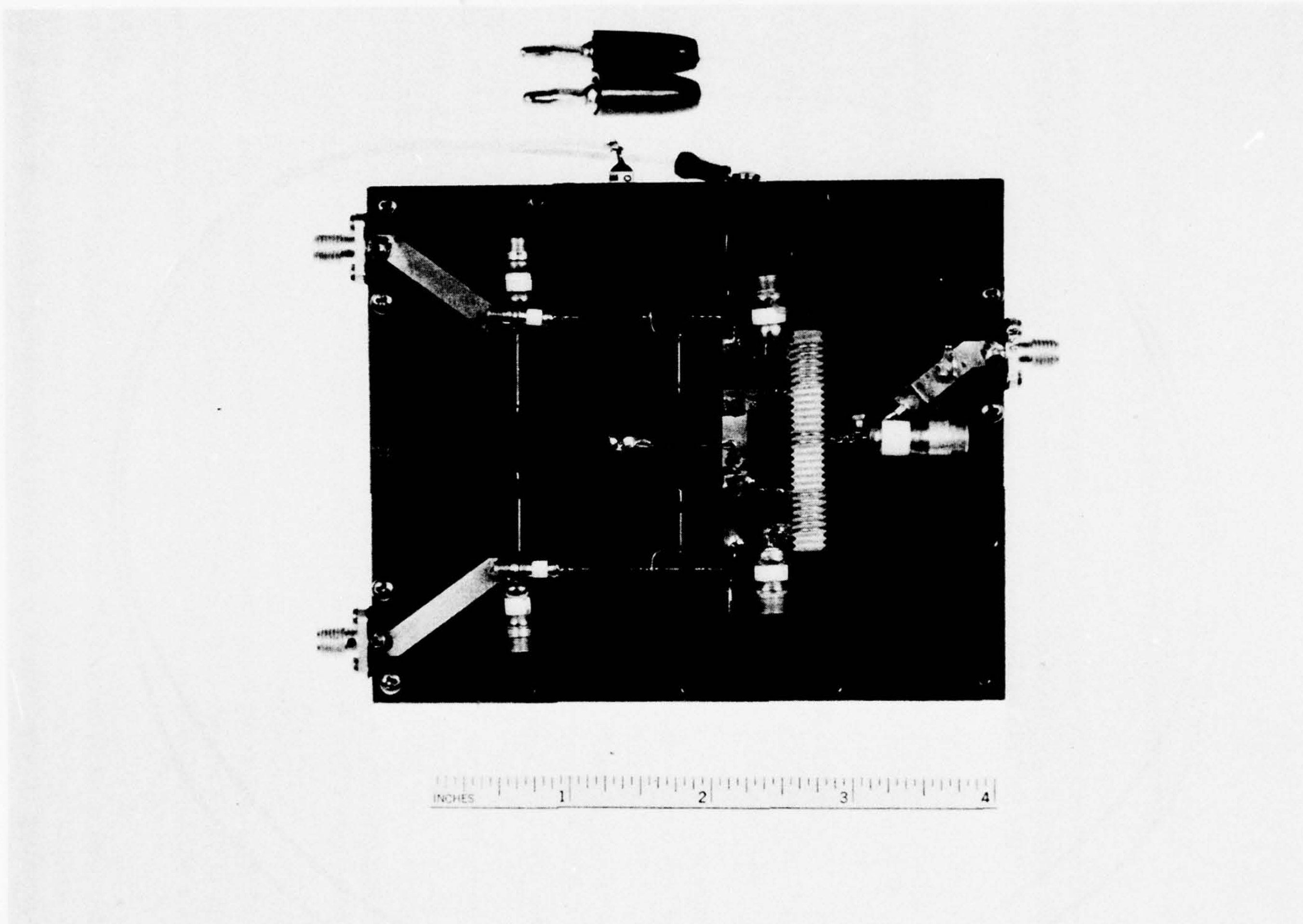


Figure 13. Experimental Model of 500- to 1000-MHz FET Mixer.

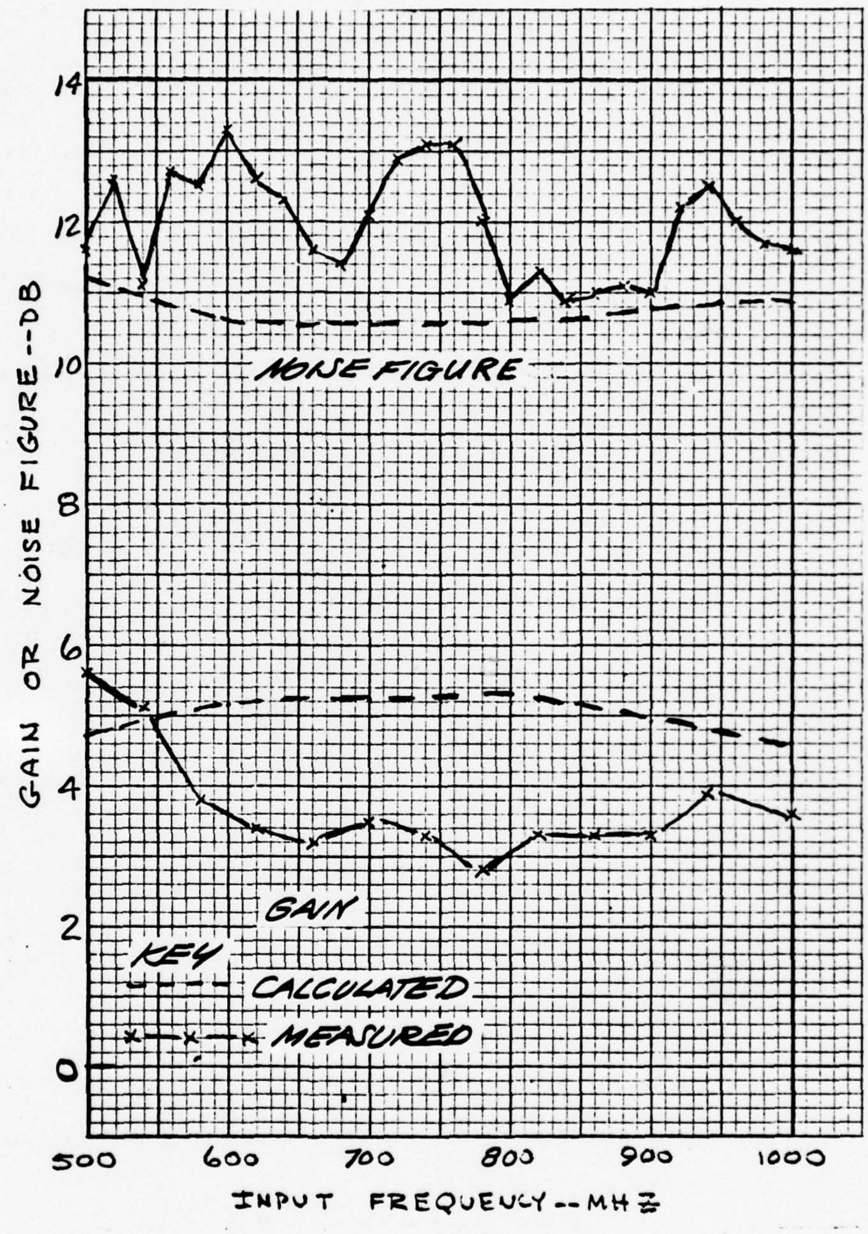


Figure 14. Calculated and Measured Noise Figure for a UHF FET Mixer.

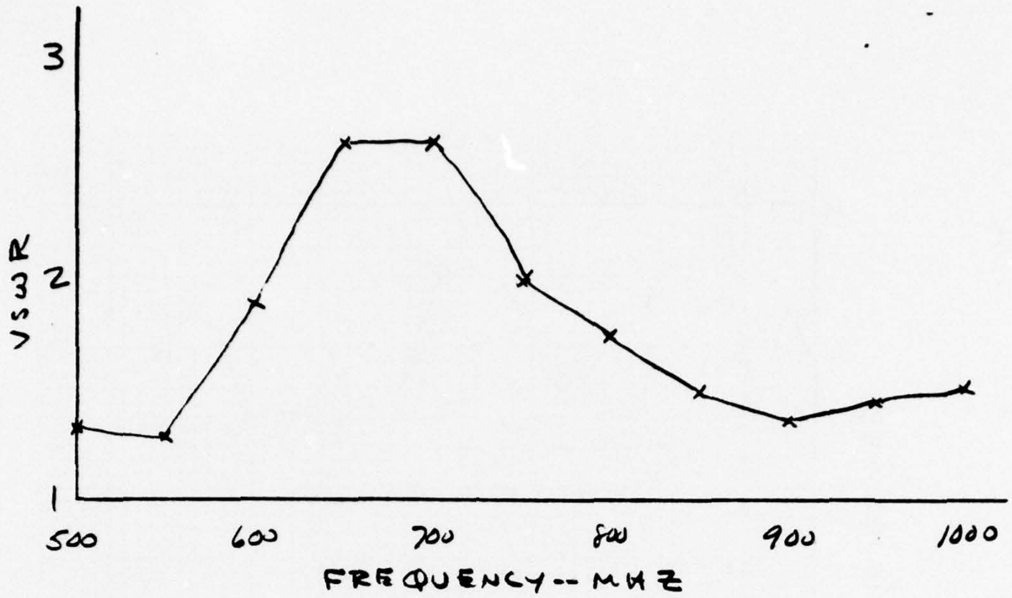


Figure 15. Input VSWR for a UHF FET Mixer.

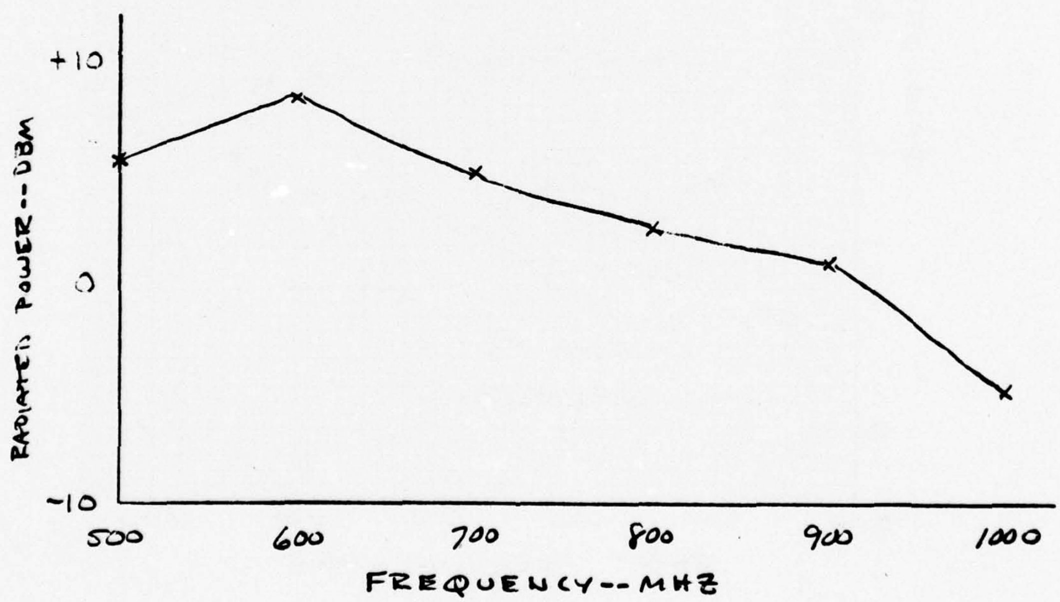


Figure 16. Local Oscillator Radiation for a UHF FET Mixer.

TABLE 1. CALCULATED AND MEASURED 500- TO 1000-MHz FIELD-EFFECT-TRANSISTOR MIXER PERFORMANCE PARAMETERS COMPARED WITH DOUBLY BALANCED MIXER PERFORMANCE.

Parameter	FET Mixer		Diode Mixer *
	Calculated	Measured	(Relcom M1A)
Input for 3-dB gain compression	+ 3.6 dBm	+ 6 dBm	+ 5 dBm
Local oscillator drive requirement	+ 15 dBm	+ 10 dBm	+ 7 dBm
Third-order intermodulation ratio for two -10-dBm input signals	67 dB	59 dB	51 dB
Intermediate-frequency bandwidth	158-162 MHz	158-162MHz	DC-1000 MHz

* Measured under same conditions as FET mixer.

The 10-MHz mixer (Figure 17) provided test results which correlated with calculations remarkably well. The tests were run with source resistances of 1000 ohms, and 2500 ohms. Table 3 is a tabulation of the results. It should be noted that the noise figure measured was single-sideband, and included about 1 dB of loss in the input tuned circuit.

In addition, third-order intermodulation distortion of 52 dB below two -16-dBm input signals was measured. This compares with a calculated value of 47.5 dB for the same input.

TABLE 2. CALCULATED AND MEASURED 500- TO 1000-MHZ FET MIXER HARMONIC INTERMODULATION RESPONSES COMPARED WITH DOUBLY BALANCED MIXER PERFORMANCE.

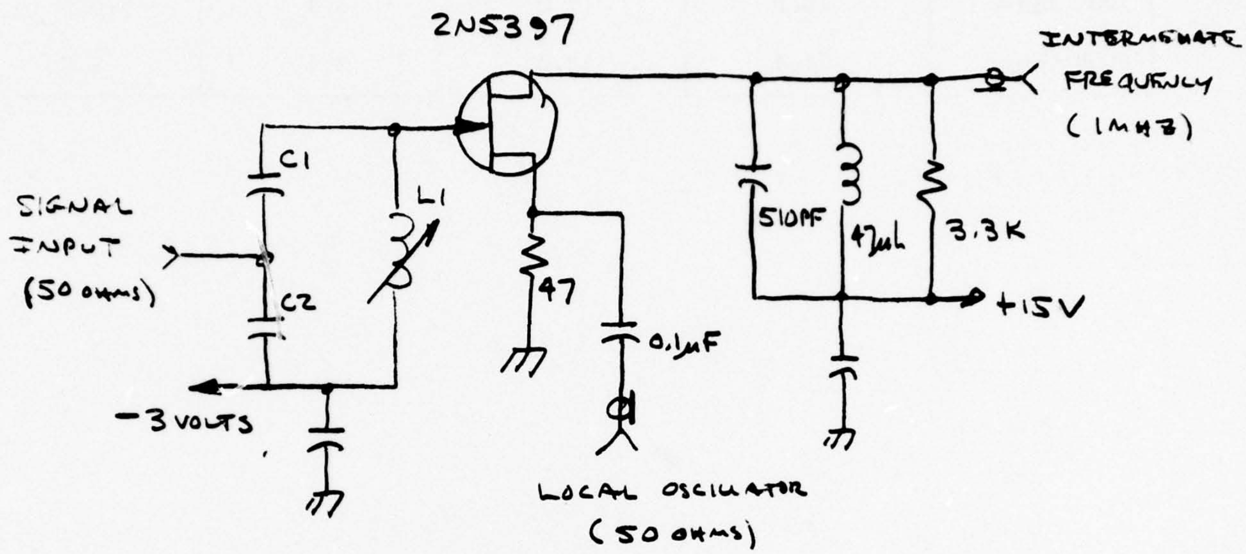
RESPONSE (M x N)*	FET Mixer		Diode Mixer** (Relcom M1A) (dB)
	Calculated (dB)	Measured (dB)	
1 x 1	0	0	0
1 x 2	37.4	36	50
2 x 1	23	8	-
2 x 2	73	40	51
2 x 3	104	65	54
3 x 2	55	42	-
3 x 3	91	65	51
3 x 4	126	84	82
4 x 3	96	92	-
4 x 4	131	96	81
4 x 5	164	115	87
5 x 4	149	92	-

* Intermediate frequency = $Mf_L \pm N_{fr}$

M refers to harmonics of the local oscillator, f_L

N refers to harmonics of the input frequency, f_R

** Measured under same conditions as FET mixer.



SOURCE RESISTANCE	C1	C2	L1
1000 OHMS	75PF	470PF	3.8μh
2500 OHMS	210PF	750PF	1.55μh

Figure 17. 10-MHz Test FET Mixer.

TABLE 3. COMPARISON OF CALCULATED AND MEASURED PERFORMANCE FOR A 10-MHZ FET MIXER -- INTERMEDIATE FREQUENCY = 1 MHZ.

Source Resistance	Gain		Noise Figure	
	Calculated (dB)	Measured (dB)	Calculated (dB)	Measured (dB)
1000 ohms	16.9	17.0	3.4	3.6
2500 ohms	21.4	22.0	2.17	2.3

Section 5

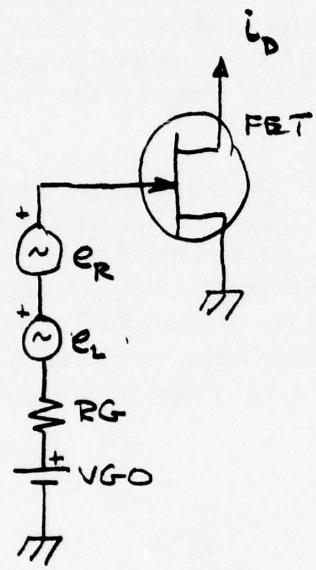
COMPUTER PROGRAM FOR CALCULATING FET MIXER PERFORMANCE

A computer program named FETMX was written in FORTRAN IV for the IBM 1130 for automating some of the calculations required to estimate FET mixer performance. The equations used in the program are based upon the model developed in Section 2 and methods outlined in References 1 and 3. The purpose of this section is to show how the program is used, and to provide flow charts of the program, a list of the variable names and their meanings, and a listing of the program.

The inputs required for the program are divided into three categories. The first consists of parameters describing the FET, which are I_{DSS}, V_p, R_S, and GAMMA. The second category provides bias point, signal and local-oscillator information. These are VGO, VLO, RGEN, and PAVS. The third is the number of harmonics desired for the local-oscillator voltage, MV (may be 1 to 5), and the highest-order term to be used for calculating harmonic intermodulation responses, NALPH (may be from 0 to 9, with 0 indicating that no harmonic intermodulation responses are desired). Figure 18 illustrates the assumed circuit (FET model is described in Section 2).

The input format for the data cards is:

Card 1:	Columns	1-10	IDSS
		11-20	VP
		21-30	RS
		31-40	GAMMA
		41-50	VGO
Card 2:	Columns	1-10	VLO
		11-20	RG
		21-30	PAVS
Card 3:	Column	1	MV
		2	NALPH



$$PAVS = \frac{EG^2}{4\ RG}$$

$$e_R = \sqrt{2}\ EG \cos \omega_R t$$

$$e_L = VLO \cos \omega_L t$$

Figure 18. Model Used for Calculating FET Mixer Parameters in FETMX.

Figures 19 and 20 are typical line-printer outputs, the first one being for a case wherein MV = 5, NALPH = 0, and the second for a case wherein MV = 5, NALPH = 8. The first example takes about 3 minutes of computer time, while the second example requires about 25 minutes of computer time.

A simplified flow chart of the computer program is provided in Figure 21, while a detailed flow chart for the calculation of harmonic responses is given in Figure 22. The calculation performed by the routine detailed in Figure 22 is

$$i = \sum_{1=1}^{NALPH} \text{ALPHA (I)} \cdot \left\{ \sum_{J=1}^{MV} V(J) \cdot \cos\left[JX(J) \cdot \omega_L + JY(J) \cdot \omega_R\right] \right\}^{I},$$

where $$i = \sum_{M=0}^{9} \sum_{N=0}^{9} \text{BETA (M, N)} \cos (M \cdot \omega_L + N \cdot \omega_R).$$

Tables 4 and 5 are lists of the variable names used in the program. Integers used as counters are not given unless they have particular significance. A listing of the program is given at the end of this section.

```
FET MIXER PERFORMANCE PARAMETERS

      IDSS  =  0.150E-01 AMPS
      VP    =  0.240E 01 VOLTS
      RS    =  0.220E 02 OHMS
      GAMMA =  0.600E 00

      VGO   =  0.192E 01 VOLTS
      VLO   =  0.190E 01 VOLTS,PEAK
      RGEN  =  0.100E 03 OHMS
      PAVS  = -0.100E 02 DBM

CALCULATED MIXER CONSTANTS--

      CONVERSION TRANSCONDUCTANCE =  0.296E-02 MHOS
      AVERAGE TRANSCONDUCTANCE    =  0.409E-02 MHOS
      NOISE TRANSCONDUCTANCE      =  0.312E-02 MHOS

      GM    =  0.450E-02 MHOS
      GC    =  0.326E-02 MHOS

      IDP   =  0.194E-02 AMPS
      VGOP  =  0.153E 01 VOLTS
      EGEN  =  0.199E 00 VOLTS,RMS

NOISE FIGURE =  0.741E 01 DB (IMAGE AND IF SHORT CIRCUITED)

THIRD ORDER INTERMODULATION RATIO IS--

      IMD3                =  0.590E 02 DB
      IMD3 (WORST CASE)   =  0.379E 02 DB
      RS FOR WORST CASE   =  0.222E 03 OHMS
```

Figure 19. Typical Output for FETMX -- Example 1.

FET MIXER PERFORMANCE PARAMETERS

```
     IDSS  =  0.143E-01 AMPS
     VP    =  0.226E 01 VOLTS
     RS    =  0.286E 02 OHMS
     GAMMA =  0.600E 00

     VGO   =  0.150E 01 VOLTS
     VLO   =  0.141E 01 VOLTS,PEAK
     RGEN  =  0.100E 03 OHMS
     PAVS  = -0.160E 02 DBM
```

CALCULATED MIXER CONSTANTS--

```
     CONVERSION TRANSCONDUCTANCE =  0.285E-02 MHOS
     AVERAGE TRANSCONDUCTANCE    =  0.439E-02 MHOS
     NOISE TRANSCONDUCTANCE      =  0.342E-02 MHOS

     GM   =  0.502E-02 MHOS
     GC   =  0.326E-02 MHOS

     IDP  =  0.225E-02 AMPS
     VGOP =  0.136E 01 VOLTS
     EGEN =  0.100E 00 VOLTS,RMS
```

NOISE FIGURE = 0.826E 01 DB (IMAGE AND IF SHORT CIRCUITED)

THIRD ORDER INTERMODULATION RATIO IS--

```
     IMD3                =  0.670E 02 DB
     IMD3 (WORST CASE)   =  0.506E 02 DB
     RS FOR WORST CASE   =  0.199E 03 OHMS
```

HARMONIC INTERMODULATION RESPONSES ARE--

HARMONICS OF FR (N) — RELATIVE LEVEL IN DB

HARMONICS OF FR (N)	0	1	2	3	4	5	6	7	8	9
9	******	******	******	******	******	******	******	******	******	******
8	-255.2	******	******	******	******	******	******	******	******	******
7	-218.2	-218.8	-235.1	-240.5	-251.7	-257.7	******	******	******	******
6	-176.0	-184.2	-195.5	-195.1	-205.1	-235.8	-218.3	-262.5	-246.8	-255.4
5	-139.9	-145.1	-171.2	-163.5	-164.4	-194.1	-183.4	-188.8	-205.1	-208.5
4	-102.2	-109.9	-130.4	-125.5	-131.2	-148.7	-147.9	-155.0	-173.7	-170.5
3	-64.1	-73.8	-104.1	-90.9	-96.2	-126.9	-115.2	-121.3	-145.9	-140.7
2	-23.7	-37.4	-73.0	-55.0	-62.2	-88.5	-81.0	-90.4	-121.3	-108.0
1	4.9	0.0	-22.9	-18.7	-27.5	-42.5	-47.9	-60.2	-83.1	-77.0
0	12.5	24.4	15.2	-29.2	-11.4	-10.3	-16.3	-33.2	-46.1	-47.8

HARMONICS OF FL (M)

IA = 8

Figure 20. Typical Output for FETMX -- Example 2.

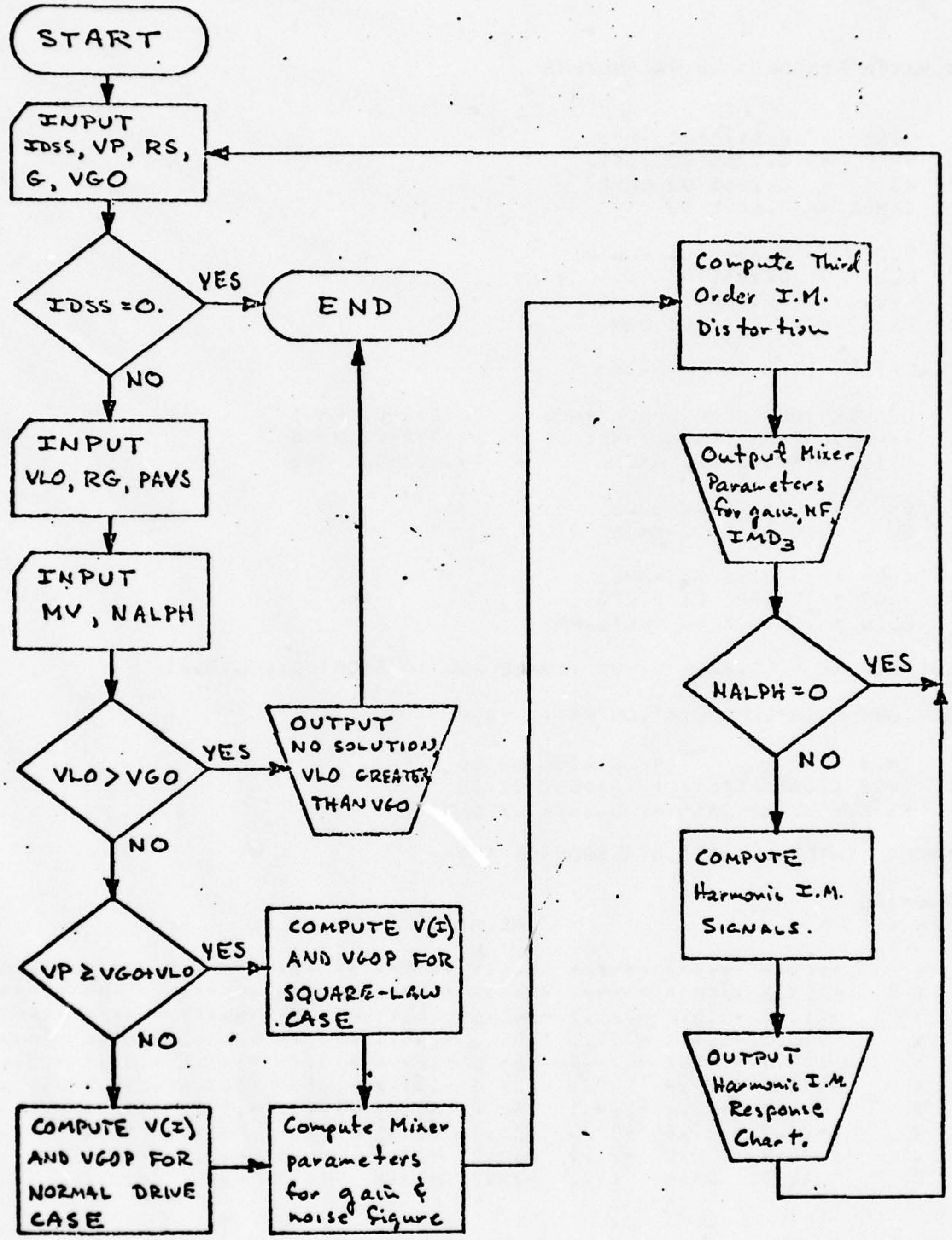
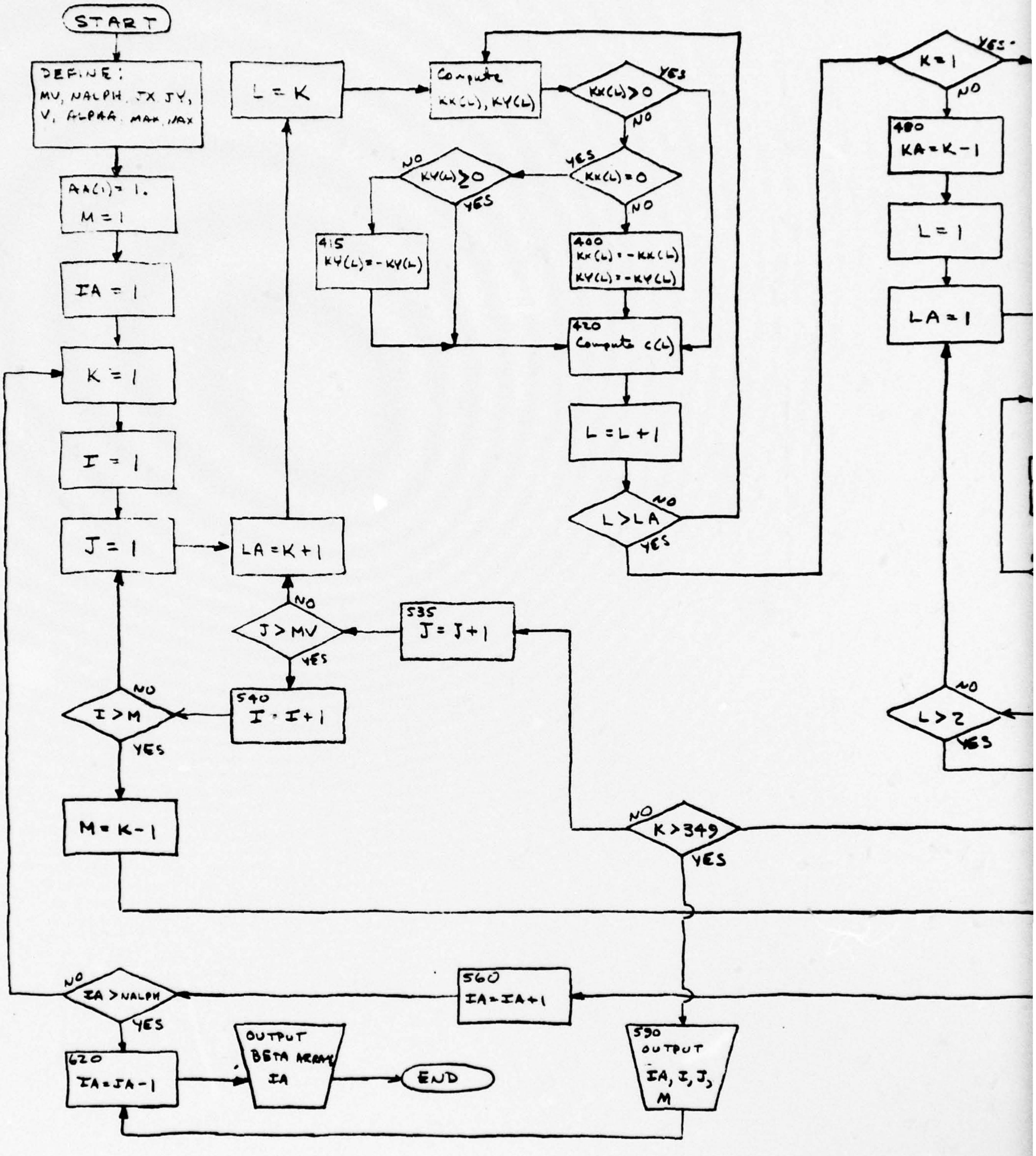


Figure 21. Program Flow Chart for FET Mixer Performance Parameters.


START
DEFINE: MV, NALPH, JX JY, V, ALPHA, MAX, NAX
AA(1) = 1.
M = 1
IA = 1
K = 1
I = 1
J = 1
LA = K + 1
L = K
Compute KX(L), KY(L)
KX(L) > 0
YES
NO
KX(L) = 0
KY(L) ≥ 0
415 KY(L) = -KY(L)
400 KX(L) = -KX(L) KY(L) = -KY(L)
420 Compute C(L)
L = L + 1
L > LA
K = 1
480 KA = K - 1
L = 1
LA = 1
L > 2
535 J = J + 1
J > MV
540 I = I + 1
I > M
M = K - 1
K > 349
560 IA = IA + 1
IA > NALPH
620 IA = IA - 1
OUTPUT BETA ARRAY IA
END
590 OUTPUT IA, I, J, M

1

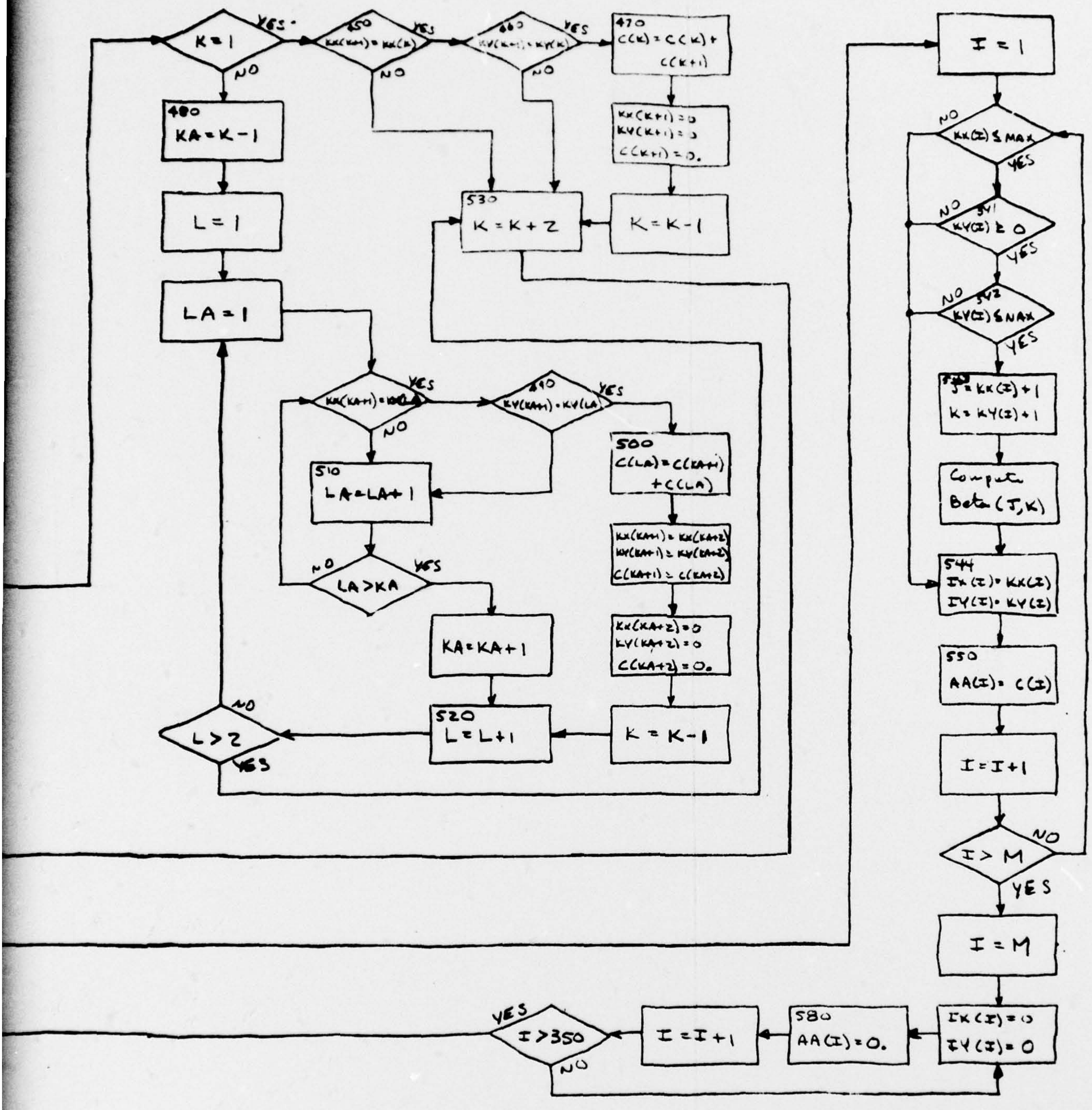


Figure 22. Flow Chart for Computation of Harmonic Intermodulation Levels.

2

TABLE 4. PRINCIPAL VARIABLE NAMES USED IN FETMX (EXCLUSIVE OF THOSE USED FOR CALCULATION OF HARMONIC INTERMODULATION RESPONSES).

IDSS	magnitude of drain current for $V_{gs} = 0$ (amperes)
VP	magnitude of pinch-off voltage (volts)
RS	bulk resistance in series with source lead (ohms)
G	gamma, a factor accounting for thermal noise modulation of channel width (usually taken as 0.6)
VGO	magnitude of no-signal gate-to-source voltage (volts)
VLO	peak local oscillator voltage (volts)
RG	generator resistance (ohms)
PAVS	available power from generator (dBm)
VGOP	effective gate-to-source voltage (volts)
GL	conversion transconductance (mhos)
GM	average transconductance (mhos)
GN2	equivalent noise transconductance (mhos)
EG	rms generator open-circuit voltage (volts)
F	noise factor for a common-source FET mixer with image frequency and intermediate frequency short-circuited at the mixer input.
IMD3	third-order intermodulation-distortion ratio.

TABLE 5. PRINCIPAL VARIABLE NAMES USED IN FETMX FOR CALCULATION OF HARMONIC INTERMODULATION RESPONSES.

JX(J)	harmonic of W_L for the Jth term in the multiplier
JY(J)	harmonic of W_R for the Jth term in the multiplier
V(J)	coefficient for the Jth term in the multiplier
ALPHA (IA)	IA-th coefficient of the power series expansion of the FET transfer characteristic
IX(I)	harmonic of W_L for the I-th term in the multiplicand
IY(I)	harmonic of W_R for the I-th term in the multiplicand
AA(I)	magnitude of the I-th term in the multiplicand
KX(K)	harmonic of W_L for the K-th term in the product
KY(K)	harmonic of W_R for the K-th term in the product
C(K)	magnitude of the K-th term in the product

COMPUTER PROGRAM LISTING

```
LOG DRIVE    CART SPEC    CART AVAIL   PHY DRIVE
  0000         0001          0001         0000

V2 M06   ACTUAL  8K  CONFIG  8K

// FOR
*LIST SOURCE PROGRAM
*ONE WORD INTEGERS
*IOCS(1132 PRINTER,CARD)
*NAME FETMX

      REAL IDSS,NF,IDP,IMD3,IMD3M
      DIMENSION V(6),ALPHA(9),IX(352),IY(352),AA(352),KX(352),KY(352),
     1C(352),JX(6),JY(6),BETA(10,10)
      DATA IX,IY,KX,KY,JX,JY,AA,C,BETA,V,PI/1408*0,1,11*0,810*0.,3.1416/

C INPUT IDSS,VP,RS,G,VGO

  100 READ(2,1)IDSS,VP,RS,G,VGO
      IF(IDSS)120,110,120
  110 WRITE(3,22)
      CALL EXIT

C INPUT VLO,RG,PAVS

  120 READ(2,1)VLO,RG,PAVS

C INPUT MV,NALPH

      READ(2,3)MV,NALPH

      IF(VLO-VGO)150,150,240
  240 WRITE(3,2)
      GO TO 110

  150 IF(VP-VGO-VLO)170,160,160

C COMPUTE V(I) AND VGOP FOR SQUARE-LAW CASE

  160 V(1)=VLO
      VGOP=VP*(1.-SQRT(1.-2.*VGO/VP+(VGO/VP)**2+(VLO/(2.*VP))**2))
      GO TO 230

C COMPUTE V(I) AND VGOP FOR NORMAL DRIVE CASE

  170 X=(VGO-VP)/VLO
      Y=SQRT(1.-X**2)
      IF(X)210,190,180
  190 A=PI/2.
      GO TO 200
  180 A=ATAN(Y/X)
      GO TO 200
  210 A=ATAN(Y/X)+PI
  200 XO=VLO-VGO+VP
```

```
      V(1)=-XO*(A-0.5*SIN(2.*A))/(PI*(1.-COS(A)))
      DO 220 I=2,MV
      JX(I)=I
  220 V(I)=-XO*(SIN((I-1)*A)/(I-1)-SIN((I+1)*A)/(I+1))/(I*PI*(1.-COS(A))
     1)
      VGOP=VP-XO*(SIN(A)-A*COS(A))/(PI*(1.-COS(A)))

C COMPUTE MIXER PARAMETERS FOR GAIN AND NOISE FIGURE

  230 GC=IDSS*ABS(V(1))/VP**2
      GM=2.*IDSS*(1.-VGOP/VP)/VP
      GN1=G*GM
      GN2=GN1+G*4*IDSS*(V(1)/(4.*VP)*(1.+VGOP/(2.*VP)+V(2)/(4.*VP)+
     1V(2)*V(3)/(4.*V(1)*VP)+V(3)*V(4)/(4.*V(1)*VP)))**2/VP
      EG=SQRT(0.004*RG*EXP(PAVS/4.3429))
      VIN=EG*SQRT(2.)
      F=1.+2.*GN1*RS**2/RG+2.*RS/RG+(GN2+RS*GM**2)*(1.+GM*RS)**2/
     1(RG*GC**2)
      NF=4.3429*ALOG(F)
      IDP=IDSS*(1.-VGOP/VP)**2
      GCP=GC/(1.+GM*RS)
      GMP=GM/(1.+GM*RS)

C COMPUTE THIRD ORDER INTERMODULATION DISTORTION

      IMD3=8.6858*ALOG(0.133*(VP/VIN)**2*(SQRT(VP/(RS*IDSS))+2.*
     1SQRT(RS*IDP))**4)
      IMD3M=8.6858*ALOG(8.5*(VP/VIN)**2*IDP/IDSS)
      RSM=0.5*VP/SQRT(IDSS*IDP)

C OUTPUT FET MODEL PARAMETERS FOR GAIN, NF,IMD3

      WRITE(3,4)
      WRITE(3,5)IDSS,VP,RS,G,VGO,VLO,RG,PAVS
      WRITE(3,6)GCP,GMP,GN2,GM,GC,IDP,VGOP,EG
      WRITE(3,7)NF,IMD3,IMD3M,RSM

      IF(NALPH)250,100,250

C COMPUTE HARMONIC INTERMODULATION SIGNALS

  250 Q=2.*RS*SQRT(IDSS*IDP)/VP
      ALPHA(1)=-2.*SQRT(IDSS*IDP)/(VP*(1+Q))
      XA=1.
      DO 300 I=2,NALPH
      XA=(2.*I-3.)*XA/I
  300 ALPHA(I)=XA*(2.*IDSS)**(I-1)*RS**(I-2)/(VP**(2*I-2)*(1+Q)**(2*I-1)
     1)
      MV=MV+1
      V(MV)=VIN
      JY(MV)=1

C MULTIPLY TWO SERIES AND STORE COEFFICIENTS IN BETA

      AA(1)=1.
```

BEST AVAILABLE COPY

```
      M=1
      DO 560 IA=1,NALPH
      K=1
      DO 540 I=1,M
      DO 535 J=1,MV
      LA=K+1
      DO 420 L=K,LA
      KX(L)=IX(I)+(2*(LA-L)-1)*JX(J)
      KY(L)=IY(I)+(2*(LA-L)-1)*JY(J)
      IF(KX(L))400,410,420
  400 KX(L)=-KX(L)
      KY(L)=-KY(L)
      GO TO 420
  410 IF(KY(L))415,420,420
  415 KY(L)=-KY(L)
  420 C(L)=AA(I)*V(J)/2.

C ROUTINE TO ELIMINATE IDENTICAL TERMS

      IF(K-1)450,450,480
  450 IF(KX(K+1)-KX(K))530,460,530
  460 IF(KY(K+1)-KY(K))530,470,530
  470 C(K)=C(K)+C(K+1)
      KX(K+1)=0
      KY(K+1)=0
      C(K+1)=0.
      K=K-1
      GO TO 530
  480 KA=K-1
      DO 520 L=1,2
      DO 510 LA=1,KA
      IF(KX(KA+1)-KX(LA))510,490,510
  490 IF(KY(KA+1)-KY(LA))510,500,510
  510 CONTINUE
      KA=KA+1
      GO TO 520
  500 C(LA)=C(KA+1)+C(LA)
      KX(KA+1)=KX(KA+2)
      KY(KA+1)=KY(KA+2)
      C(KA+1)=C(KA+2)
      KX(KA+2)=0
      KY(KA+2)=0
      C(KA+2)=0.
      K=K-1
  520 CONTINUE
  530 K=K+2

      IF(K-351)535,535,590
  590 WRITE(3,8)IA,I,J,M
      GO TO 620
  535 CONTINUE

  540 CONTINUE
      M=K-1

C M=TOTAL NUMBER OF TERMS STORED IN (KX,KY,C)
```

```
C MAX=9, NAX=9

      DO 550 I=1,M
      IF(KX(I)-9)541,541,544
  541 IF(KY(I))544,542,542
  542 IF(KY(I)-9)543,543,544
  543 J=KX(I)+1
      K=KY(I)+1
      BETA(J,K)=BETA(J,K)+C(I)*ALPHA(IA)
  544 IX(I)=KX(I)
      IY(I)=KY(I)
  550 AA(I)=C(I)
      LA=M+1
      DO 580 I=LA,350
      IX(I)=0
      IY(I)=0
  580 AA(I)=0.
  560 CONTINUE

  620 IA=IA-1
      BB=BETA(2,2)
      DO 610 J=1,10
      DO 600 I=1,10
  600 BETA(I,J)=8.6858*ALOG(BETA(I,J)/BB)
  610 CONTINUE

C OUTPUT HARMONIC INTERMODULATION RESPONSE CHART

      WRITE(3,9)
      DO 330 I=1,10
      K=10-I
  330 WRITE(3,10)K,(BETA(J,K+1),J=1,10)
      WRITE(3,21)IA
      GO TO 100

C FORMAT STATEMENTS

    1 FORMAT(8F10.4)
    2 FORMAT('1NO SOLUTION, VLO-VGO GREATER THAN ZERO')
    3 FORMAT(2I1)
    4 FORMAT('1FET MIXER PERFORMANCE PARAMETERS'//)
    5 FORMAT(6X,'IDSS  = ',E10.3,' AMPS'/6X,'VP    = ',E10.3,' VOLTS'/6X
     1,'RS    = ',E10.3,' OHMS'/6X,'GAMMA = ',E10.3,//6X,'VGO   = ',E10.
     23,' VOLTS'/6X,'VLO   = ',E10.3,' VOLTS,PEAK'/6X,'RGEN  = ',E10.3,'
     3 OHMS'/6X,'PAVS  = ',E10.3,' DBM')
    6 FORMAT('OCALCULATED MIXER CONSTANTS--'//6X,'CONVERSION TRANSCONDUC
     1TANCE = ',E10.3,' MHOS'/6X,'AVERAGE TRANSCONDUCTANCE     = ',E10.3,
     2                                   ' MHOS'/6X,'NOISE TRANSCONDUCT
     3ANCE       = ',E10.3,' MHOS'//6X,'GM    = ',E10.3,' MHOS'/6X,'GC    =
     4 ',E10.3,' MHOS'//6X,'IDP   = ',E10.3,' AMPS'/6X,'VGOP = ',E10.3,'
     5VOLTS'/6X,'EGEN = ',E10.3,' VOLTS,RMS')
    7 FORMAT('ONOISE FIGURE = ',E10.3,' DB (IMAGE AND IF SHORT CIRCUITED
     1)'//' THIRD ORDER INTERMODULATION RATIO IS--'//6X,'IMD3',14X,'= ',
     2E10.3,' DB'/6X,'IMD3 (WORST CASE) = ',E10.3,' DB'/6X,'RS FOR WORST
     3 CASE = ',E10.3,' OHMS')
    8 FORMAT('OK OVER- IA,I,J,M = ',4I4)
```

```
  9 FORMAT('0HARMONIC INTERMODULATION RESPONSES ARE--'/'0HARMONICS'/'
   1OF FR (N)',24X,'RELATIVE LEVEL IN DB'/)
 10 FORMAT(5X,I1,4X,10(1X,F6.1))
 21 FORMAT(/13X,'0',6X,'1',6X,'2',6X,'3',6X,'4',6X,'5',6X,'6',6X,'7',6
   1X,'8',6X,'9'/32X,'HARMONICS OF FL (M)'/' IA = ',I2)
 22 FORMAT(3('1'/))

    END

FEATURES SUPPORTED
 ONE WORD INTEGERS
 IOCS

CORE REQUIREMENTS FOR FETMX
 COMMON       0  VARIABLES   3154  PROGRAM   2316

 END OF COMPILATION

// DUP

*STORE       WS  UA  FETMX
CART ID 0001   DB ADDR  1D45   DB CNT   014C

// XEQ FETMX
```

Section 6

SUMMARY

A study has been made of theoretical and experimental performance of a 90-degree-hybrid-coupled FET mixer covering the 500- to 1000-MHz frequency range, using microstrip techniques. The theoretical work completed includes methods for calculating FET mixer noise figure, third-order intermodulation, and harmonic intermodulation signals (single-frequency spurious responses). The methods for calculating third-order intermodulation and harmonic intermodulation represent extension of work already found in the literature.[1] The work on mixer noise figures takes into account noise contributions at the input, image and intermediate frequencies. It includes noise caused by the bulk resistance in series with the drain and source leads, and channel-width modulation arising from thermal noise in the conducting channel of the FET. This effect has been well described in the literature, but no description for mixers has been located. Because the noise source related to channel-width modulation is a function of the gate-to-source voltage, it is time varying at the local-oscillator frequency and must be handled appropriately. This effect was included as a part of the analysis. The test results showed that the noise-figure calculations were within 1 or 2 dB of the measured values, but that larger differences were encountered in the intermodulation measurements, indicating the need for additional work on the model for intermodulation performance.

Test data from the experimental model of the mixer indicated a frequency coverage of from 500 to 1000 MHz, a noise figure of from 10.9 to 13.3 dB, gain of 2.8 to 5.6 dB, and a 3-dB-gain-compression level of +6-dBm input. Third-order intermodulation performance of 59 dB below two -10-dBm input signal levels was also measured. The intermediate frequency is 160 MHz with a bandwidth of 4 MHz.

In addition, a computer program named FETMX, written in FORTRAN IV for the IBM 1130 and automating the calculations required to estimate FET mixer performance, is described. It computes mixer constants such as conversion transconductance, single-sideband noise figure, third-order intermodulation distortion, and harmonic intermodulation responses, including those caused by curvature up through the ninth order.

During the course of this study, and as a result of work on other FET mixer designs, coupled with extensive experience with doubly balanced diode mixers, some general qualitative judgments about the relative merits of the two types of mixers have been developed. These conclusions can be listed as follows:

a. Third-order intermodulation distortion in an FET mixer is almost always greater than in a doubly balanced diode mixer for the same input power level.

b. Low-order harmonic intermodulation responses are lower-level for diode mixers, and high-order harmonic intermodulation responses (say 7th order or higher) are lower for FET mixers, in comparable situations.

c. At frequencies below about 300 MHz, FET mixers can be built with lower noise figures, using presently available devices. For example, a 300-MHz FET mixer should realize a less than 4-dB noise figure and a 10-dB gain, while a typical diode mixer would have a 6-dB noise figure and 6-dB loss.

d. In fixed-frequency applications, an FET mixer is quite often a more satisfactory solution, since it combines mixing action and gain into one device.

e. The wide variety of FET's presently available on the market creates a flexibility in design which is not possible with diode mixers.

f. The fact that FET mixers have gain is very significant and should not be overlooked when considering a mixer for a given situation. For example, in most receiver applications, a diode mixer followed by a standard IF system will require a preamplifier before the mixer, to achieve a desired overall noise figure; while an FET mixer with gain may not need preamplification. Thus, the signal levels at the FET mixer input will be lower than at the diode mixer input, with consequent improvement in spurious-signal rejection ratios.

Section 7

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13. ABSTRACT

A study is presented of theoretical and experimental performance of a 90-degree-hybrid-coupled FET mixer covering 500-1000 MHz. The theoretical work completed includes methods for calculating FET mixer gain, noise figure, third-order intermodulation, and harmonic intermodulation products (single-frequency spurious responses). A model suitable for use in mixer design is also described.

Test data from the experimental model of the mixer, over the frequency range of 500 to 1000 MHz, indicated a noise figure of 10.9 to 13.3 dB, gain of 2.8 to 5.6 dB, and a 3-dB-gain-compression level of +6 dBm at the input. The intermediate frequency was 160 MHz, with a bandwidth of 4 MHz. Good correlation between calculated and measured performance was obtained. In addition, there is described a computer program, written in FORTRAN IV for the IBM 1130 and automating the calculations required to estimate FET mixer performance, in which curvative effects up through the ninth order are included.

Some comparison is made of the test results with typical hot-carrier diode mixers which, in the 500- to 1000-MHz range, are comparable except for their conversion losses.

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Intermodulation Products		1				
Hot-Carrier Diode Mixer		1				
Doubly Balanced Mixer		1				